STEADY STATE OPERATING CHARACTERISTICS OF HYBRID LOADED RESONANT POWER CONVERTER

TAUFIK
Department of Electrical Engineering
California Polytechnic State University
San Luis Obispo, CA 93407

INTRODUCTION

Design of modern power electronic systems has become demanding for high quality, small, lightweight, reliable, and efficient power processors. These objectives can be accomplished by the use of a higher switching frequency. The higher the operating switching frequency, the smaller and lighter the transformer, filter inductors, and capacitors. In addition, dynamic characteristics of converters improve with increasing operating frequencies. The bandwidth of a control loop is usually determined by the corner frequency of the output filter. Therefore, high switching frequencies allow for achieving a faster dynamic response to rapid changes in the load current and/or the input voltage. As an example, the choice of a higher switching frequency in a switch mode power supply, greatly helps to make the unit smaller and more compact by reducing the size of the filter elements. Therefore, in the design of switch mode power supplies, the aim is to use as high a switching frequency as possible. The main obstacle in increasing the frequency to much larger values is the increased switching power loss in the switching device. The switching power loss is defined to be the average of energy being dissipated at each turn on and turn off transition. It has been known that this switching power is directly proportional to the switching frequency. Hence, increasing the switching frequency would also mean increasing the switching loss. This causes not only higher temperature rise, but also lower the overall efficiency of power conversion. The strategy is then to come up with converter topology whose switching loss does not strictly depend on the switching frequency. In other words, the topology will eliminate or significantly minimize switching loss while at the same time implementing higher switching frequency, and thus attaining low component stress, improved reliability, low electromagnetic interference, low size and weight of components and higher power conversion efficiency. Resonant converter topology is well suited for this purpose.

The concept of resonant topology is realized from the fact that power dissipation during switching transition is the product of the instantaneous current and voltage. If either the current through the switch or the voltage across the switch is zero, or have a low value during the switching transition, then the power dissipation will be zero or low. In the switch mode operation where resonant topology is not used, the switches are required to turn on and turn off the entire load current during each switching. Therefore, the switches are subjected to high switching stresses, high EMI, and high switching power loss that increases linearly with the switching frequency. These problems are avoided by the use of resonant topology in that the topology provides either zero voltage switching, zero current switching, or both during switching transitions. Zero voltage switching occurs when the switch is operated while the voltage across it is zero. Likewise, zero current switching is obtained when the switch is either turned on or off while the current through it is zero.

There have been many resonant topologies proposed in recent years, and one that will be discussed in this paper is called the loaded resonant converter. In this category, an L-C tank is used that is made to resonate, thereby causing both the voltage and current go to zero naturally. There are several types of loaded resonant converter [1], two of which are called the Series Loaded and Parallel Loaded resonant converters. In the Series type, the load appears as in series to the L-C tank, while in the Parallel type the load appears as in parallel to the L-C tank. Each topology has both advantages and disadvantages. Therefore, another type has been proposed that combines the two types together, hence the name Hybrid Loaded. In this paper, steady state characteristics of Hybrid Loaded resonant converter will be presented. The knowledge of such characteristics is extremely important to realize the benefits of using the topology under different conduction modes. Parameters such as the steady state average current and the steady state average voltage at different switching frequency can be easily determined from this study, and therefore will aid us in designing a dc-dc converter circuit using hybrid topology. Computer simulations using Orcad Pspice v9.1 will accompany the discussion to help us analyze the steady state waveforms possessed by the converter under different conduction modes.

LOADED-RESONANT CONVERTERS

The term loaded refers to an LC tank that produces oscillating voltage and current applied to load. Therefore, in this type of converter, the converter switches can be switched at zero voltage and/or zero current. The LC tank

Itself may be connected in series, parallel, or a combination of both. In series loaded resonant converter (SLR), its load is in series with the resonant circuit elements. Likewise, in parallel loaded resonant converter (PLR) its load is in parallel with the resonant circuit elements. Moreover, both SLR and PLR have several modes of operation, including two categories of Continuous Conduction Mode (CCM), and Discontinuous Conduction Mode (DCM). When operated in DCM, the series loaded resonant converter acts as a current source which implies an inherent current limiting under short circuit conditions [2]. To the parallel loaded resonant converter, the DCM operation provides a voltage source characteristics, which implies load voltage regulation. In order to benefit from the advantageous properties of both the SLR and the PLR converters, the hybrid loaded resonant converter topology was proposed. The hybrid topology consists of a series-resonant circuit as shown in Figure 1 in a half bridge configuration, but the load is connected in parallel with only part of the capacitance, for example, one-third of the total capacitance, and the other two thirds of the capacitance appears in series.

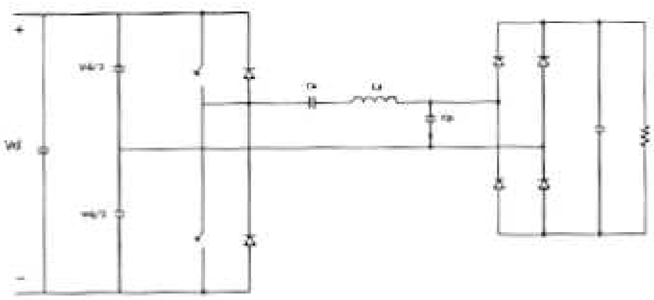

Figure 1. Hybrid Loaded Resonant Converter

STEADY STATE ANALYSIS OF LOADED-RESONANT CONVERTERS

To begin the analysis, it is important to realize that the switches produce a square-wave voltage at the input terminals of the resonant tank. Let us call this voltage v_{ir}. The output voltage of the LC tank, denoted as v_{or}, is ideally a sinusoidal waveform at the fundamental frequency of the input square wave. This output voltage is also the input voltage to the bridge rectifier. It is also assumed that the configuration as shown in Figure 1 may include an output inductor that produces a ripple-free current causing the input current to the bridge rectifier, let us call it i_{outr}, to be a square wave. The output voltage across the load resistor is V_o. The analysis also considers the fundamental frequency of the Fourier series for the voltages and currents. Therefore, the amplitudes of the fundamental frequencies of the square waves v_{ir} and v_{or} are:

$$V_{ir1} = \frac{2V_d}{\pi} \tag{1}$$

$$V_{or1} = \frac{V_o \pi}{2} \tag{2}$$

The relationship between input and output is estimated from ac analysis of the circuit for these fundamental components of the square waves. The ac equivalent circuit for hybrid topology is shown in Figure 2.

Figure 2. Equivalent ac circuit

Where R_e is the value of the output resistance based on the ratio of voltage to current at the output and is found to be:

$$R_e = \frac{\pi^2}{8} R_{Load} \tag{3}$$

By using phasor analysis on the equivalent ac circuit, it can be easily proven that:

$$\frac{v_{or1}}{v_{ir1}} = \left| \frac{1}{1 - \frac{X_{Ls}}{X_{Cp}} + \frac{X_{Cs}}{X_{Cp}} + j\left(\frac{X_{Ls}}{R_e} - \frac{X_{Cs}}{R_e}\right)} \right| \tag{4}$$

with the reactances at the switching frequencies are:

$$X_{Cs} = \frac{1}{\omega_s C_s} \tag{5}$$

$$X_{Cp} = \frac{1}{\omega_s C_p} \tag{6}$$

$$X_{Ls} = \omega_s L_s \tag{7}$$

Then, by using equations (1) and (2), equation (4) can be arranged to yield:

$$\frac{v_o}{v_d} = \frac{4}{\pi^2} \left| \frac{1}{1 - \frac{X_{Ls}}{X_{Cp}} + \frac{X_{Cs}}{X_{Cp}} + j\left(\frac{X_{Ls}}{R_e} - \frac{X_{Cs}}{R_e}\right)} \right| \tag{8}$$

Or, in terms of the switching frequency ω_s:

$$\frac{v_o}{v_d} = \frac{4}{\pi^2\sqrt{\left(1+\frac{C_p}{C_s}-\omega_s^2 L_s C_p\right)^2+\left(\frac{\omega_s L_s}{R_e}-\frac{1}{\omega_s R_e C_s}\right)^2}} \tag{9}$$

If we define a quality factor:

$$Q = \frac{\omega_o L_s}{R_L} \tag{10}$$

then, equation (9) can be rewritten as follows:

$$\frac{v_o}{v_d} = \frac{4}{\pi^2\sqrt{\left(1+\frac{C_p}{C_s}-\left[\frac{\omega_s}{\omega_o}\right]^2\right)^2+\frac{64}{\pi^2}Q^2\left(1-\frac{1}{[\omega_s/\omega_o]}\right)^2}} \tag{11}$$

If v_o/v_d is denoted as the normalized output voltage v_{oN}, and ω_s/ω_o as the normalized switching frequency ω_N, then the normalized frequency response can be plotted by using equation (11) for different values of Q. Figure 3 shows such a plot when $C_p = C_s$.

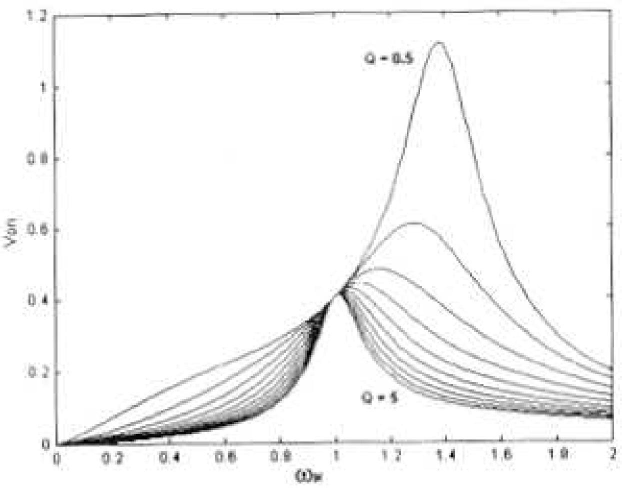

Figure 3. Normalized frequency response with C_p=C_s

The plot as depicted in Figure 3 shows that at higher Q values, the steady state frequency response merges to same values when $\omega_N < 0.5$. This interval when the switching frequency is less than half of the resonant frequency, as will be discussed later, is when the resonant converter operates in discontinuous conduction mode. Therefore, during discontinuous conduction mode, when there is a change in the load, as long as the Q value is relatively high, the output voltage v_o stays constant, i.e. the converter exhibits the output voltage regulation.

Another steady state characteristic that may be of our interest is to see what affect does the ratio of the two capacitors have on the steady state frequency response. This is significant since the purpose of having the hybrid topology is to expose only part of the capacitance in the resonant LC tank to the load. To do this, equation (10) is rearranged to yield:

$$\frac{v_o}{v_d} = \frac{4}{\pi^2\sqrt{\left(1+n-n\left[\frac{\omega_s}{\omega_o}\right]^2\right)^2+\frac{64}{\pi^2}Q^2\left(1-\frac{1}{[\omega_s/\omega_o]}\right)^2}} \tag{12}$$

where n is the ratio C_p/C_s. Figure 4 illustrates the steady state frequency response with varying ratio n, and a fixed value of Q = 4.

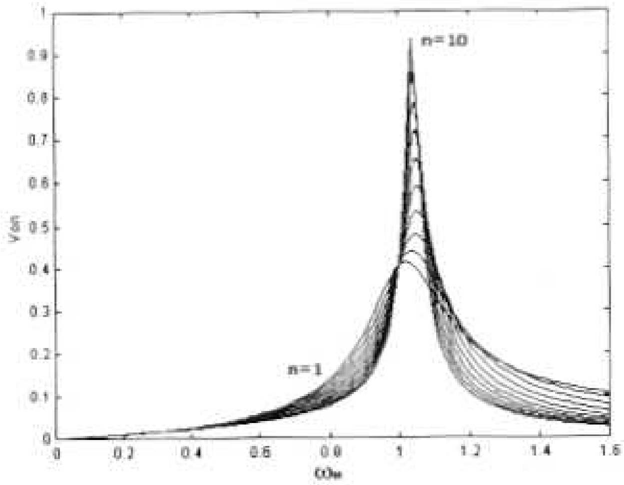

Figure 4. Normalized frequency response with Q=4

Figure 4 shows that again when the converter operates in discontinuous conduction mode, that is $\omega_N < 0.5$, all the plots from different values of n merges into the same line and hence values. This implies that at any switching frequency in the discontinuous conduction mode interval, the ratio of the resonant capacitors has no effect on the output voltage. In other words, different values of n will give the same value of output voltage v_o at any given switching frequency ω_s that is less than half of the resonant frequency ω_o. However, this not the case for other switching frequencies, as indicated by Figure 3. At about ω_N a little bit more than 1, the variation is quite noticeable that a circuit designer needs to take into consideration the effect of the different ratios n when it is

desired to operate the resonant converter at around this switching frequency.

DISCONTINUOUS CONDUCTION MODE WITH $\omega_s < \frac{1}{2}\,\omega_o$

In order to investigate the steady state waveforms under different conduction modes, Pspice computer simulation is utilized to model the resonant converter. Figure 5 and Figure 6 show the steady state waveforms of resonant inductor current and series capacitor voltage when the circuit is operated in the discontinuous conduction mode.

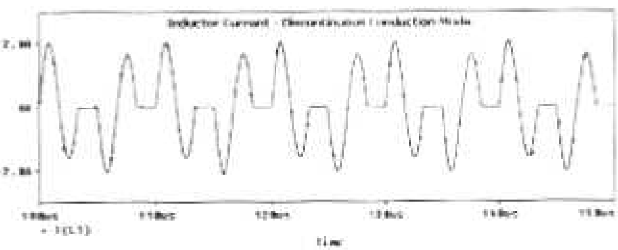

Figure 5. Steady state inductor current in DCM

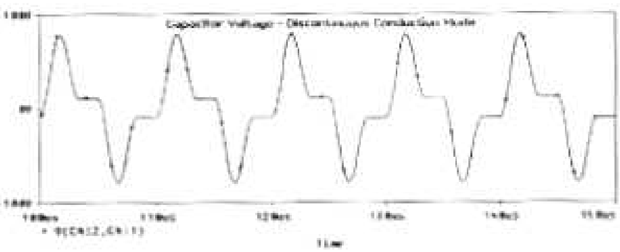

Figure 6. Steady state capacitor voltage in DCM

Since the switching frequency used is 100kHz, we can start our analysis at t = 100us. At this point, the top switch of the half bridge, denoted T_{top}, is turned on causing a zero current switching at turn on. However, since there was no current prior to t = 100us, therefore the switch is turned on at a finite voltage of $V_d/2$. Inductor current then resonates and goes to zero naturally at which point the switch T_{top} is turned off. Because of this, the switch turns off at zero current. To see whether or not the switch turns off also at zero voltage, we realize that since the current is trying to go negative after it crosses zero, then diode D_{top} which is in parallel with switch T_{top} has to carry the negative current since the switch is a unidirectional switch. Hence, since the diode is in parallel with the switch, and it is carrying the current therefore the voltage across the diode, and thus the switch T_{top} will be zero ideally, or low. The next half cycle then continues with similar operation. To summarize, in the discontinuous conduction mode, the switches turn on at zero current but not at zero voltage. Also the switches turn off naturally at zero current and at zero voltage.

CONTINUOUS CONDUCTION MODE WITH $\frac{1}{2}\,\omega_o < \omega_s < \omega_o$

In this mode, the inductor current does not remain zero for some finite amount of time as what we have seen in the case of DCM. Here, the switches are forced to turn on to carry the current flowing through one of the diodes. Figures 7 and 8 depict both steady state inductor current and capacitor voltage waveforms respectively.

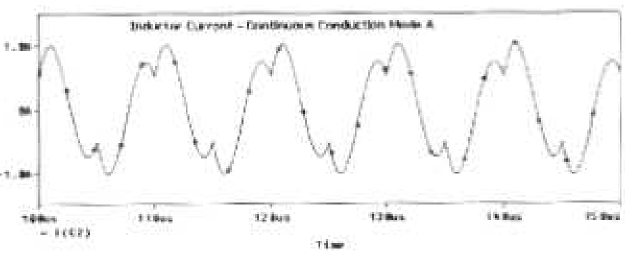

Figure 7. Steady state inductor current in DCM

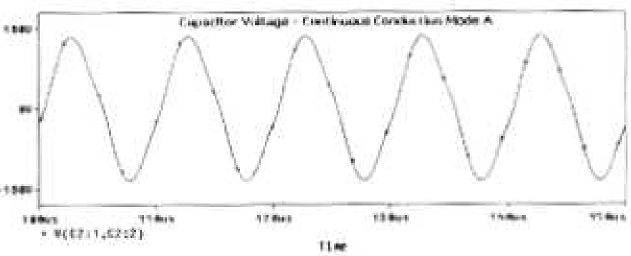

Figure 8. Steady state capacitor voltage in DCM

Taking t = 100us as the starting point, we can easily see that the switch T_{top} turns on at a finite current. Moreover, since the diode connected in parallel with the bottom switch was conducting prior to turning on T_{top}, then the switch T_{top} must turn on at also a finite voltage V_d. This results in a turn-on switching loss. As for the turn off transition, the switch T_{top} is turned off at the instant when the current resonates to zero, thus making it a zero current switching at turn off. Again, since the remaining negative current is transferred to the diode parallel to it, then the switch has zero voltage across it when it turns off. Therefore, if the resonant converter is operated under this interval, then the turn-on switching loss is expected to have an impact, while the turn-off loss is eliminated or minimized. Another disadvantage would be the choice of the diodes. Since both diodes are forced to turn off when

they are carrying a current, then the diodes must have good reverse-recovery characteristics.

CONTINUOUS CONDUCTION MODE WITH $\omega_s > \omega_o$

In this mode of operation, the switches turn on naturally at zero current, as can be easily observed from Figure 9. It is realized that diode D_{top} was conducting prior to turning on the switch T_{top}, and hence zero voltage at turn on. Consequently, the turn-on switching loss is eliminated or minimized. However, Figure 9 also shows that the switches are forced to turn off when they are still carrying some current, resulting in a turn off switching loss. This is a major disadvantage especially that the switch is turned off at close to the peak of the inductor current. Finally, the two waveforms show an improved sinusoidal quality compared to the previous two modes. Therefore, the equations that were derived previously will yield more accurate results if the resonant converter is operated under this conduction mode.

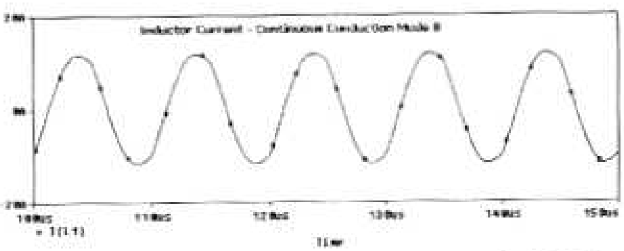

Figure 9. Steady state inductor current in DCM

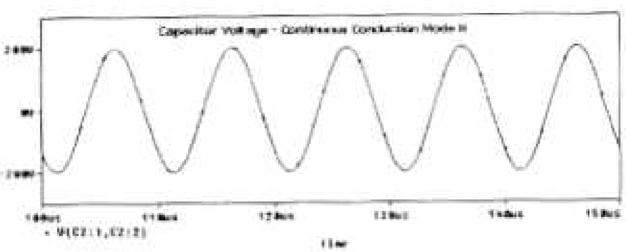

Figure 10. Steady state capacitor voltage in DCM

CONCLUSIONS

In this paper, steady state analysis of hybrid loaded resonant converter was investigated, and equations governing the steady state conditions were also derived. Using computer simulations, it was possible to observe the steady state operation of the converter under three different modes. Each mode was found to have both advantages and disadvantages in terms of the turn-on and turn-off switching losses, the choices of diodes and the switch. Knowledge of such steady state characteristics is therefore very crucial in designing the hybrid loaded resonant circuit.

REFERENCES

1. Mohan, N., Undeland, T.M. and Robbins, W.P., Power Electronics: Converters, Applications and Design, John Wiley and Sons, Inc., New York, 1995.

2. Krein, P.T., Elements of Power Electronics, Oxford University Press, Inc., Oxford, 1998.

Taufik (Member IEEE) was born in Jakarta, Indonesia. He received his B.S. degree from Northern Arizona University in 1993 and his M.S. degree from the University of Illinois at Chicago in 1995, both in Electrical Engineering. He then worked as a researcher for the Microelectronics Lab at UIC & Argonne National Lab before pursuing a doctoral degree at Cleveland State University. During the course of his study, he has been employed by Capstone Turbine Corp., Rockwell Automation, and Picker International. He received his Doctor of Engineering degree in Electrical Engineering from Cleveland State University in 1999. His research interests include Power Converter topologies, Resonant Converters, and Waveguides. Dr. Taufik is currently an Assistant Professor in Electrical Engineering at California Polytechnic State University in San Luis Obispo.